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(54) Pulse doppler radar

(57) A receiver for pulse Doppler radar equipment using p.r.f. staggering includes a digital filter arrangement for reducing clutter within a lower speed range (Sg) and a higher speed range (Sm). The arrangement includes a first digital filter (DF1) having a pass band/stop band characteristic such that the lower clutter is rejected by the lowest stop band of the first filter but the desired target echo signal(s) falls within a pass band, a circuit (HK) connected to an output of the first filter for calculating a dominating frequency component of the higher clutter (Sm) and carrying out a downwards transfer in frequency of the clutter signals recovered from the first filter so that the mean frequency of the higher clutter assumes a value within the first stop band of a second digital filter (DF2), substantially coinciding with the first stop band of the first filter in order to suppress the higher clutter.

Fig. 1

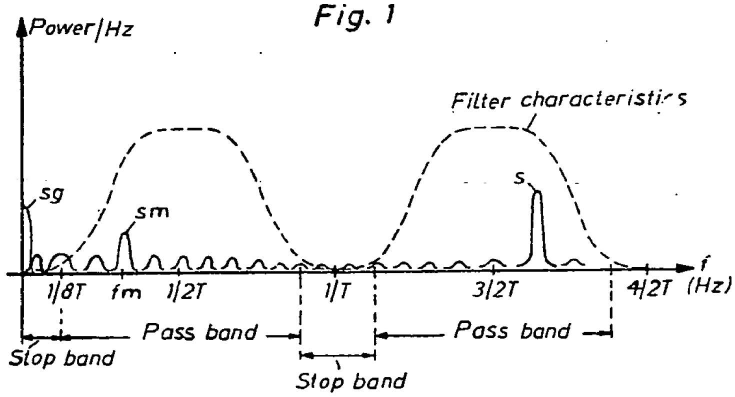

Fig. 3

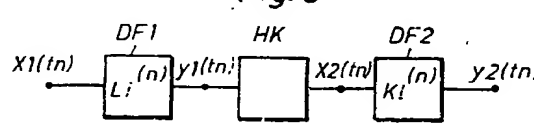

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Fig. 1

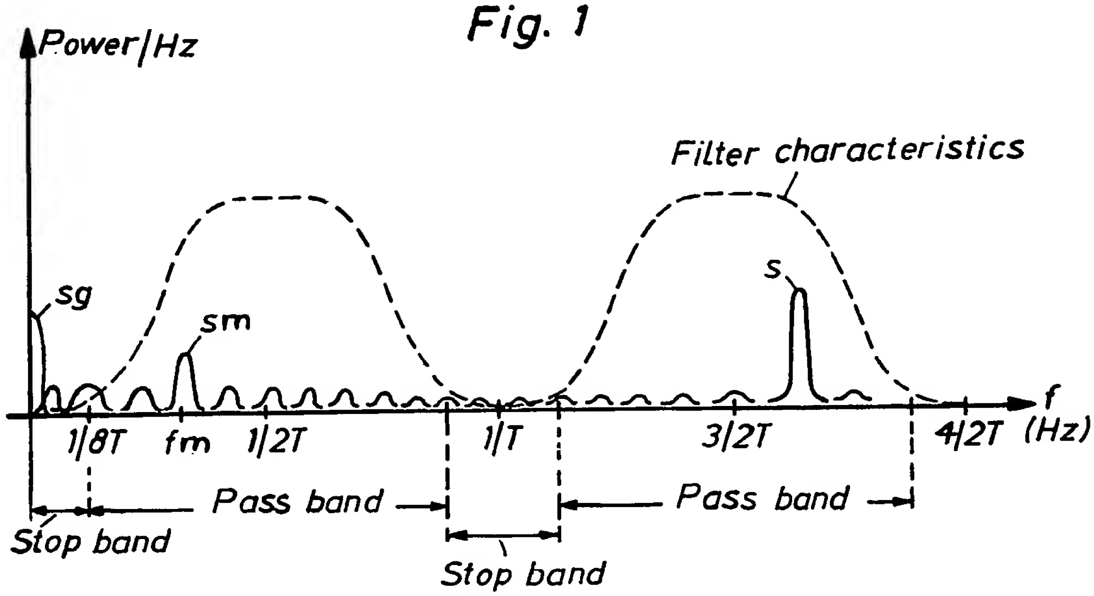

Fig. 2

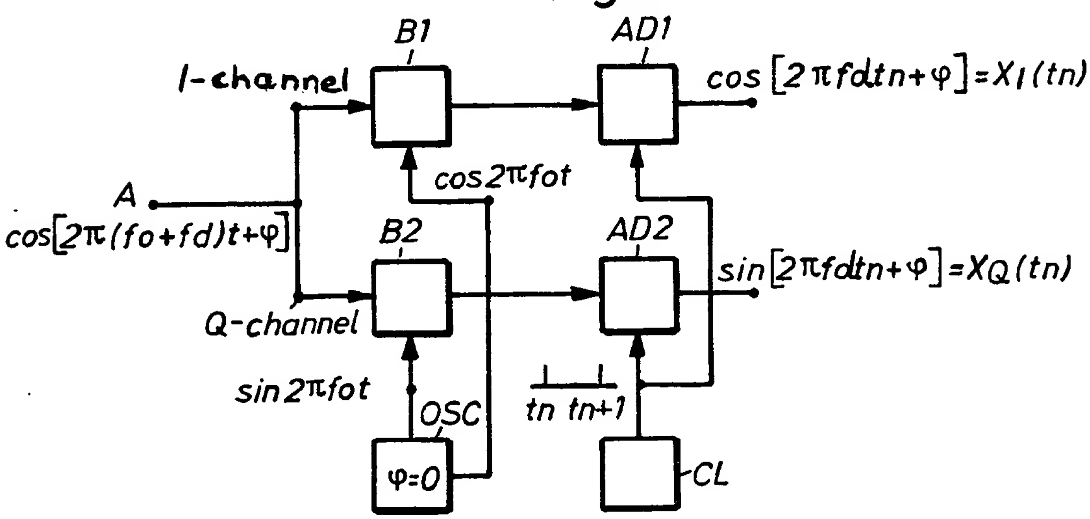

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Fig. 3

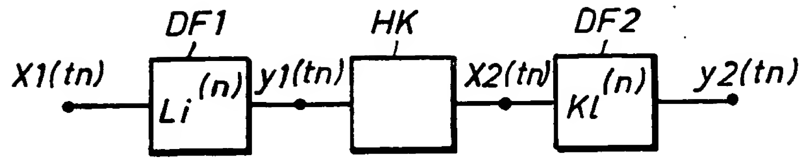


Fig. 4

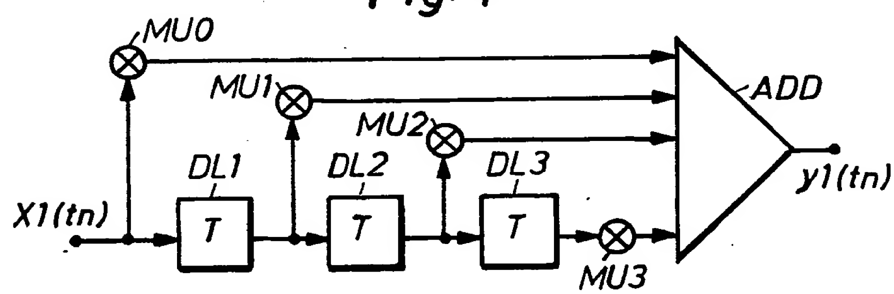
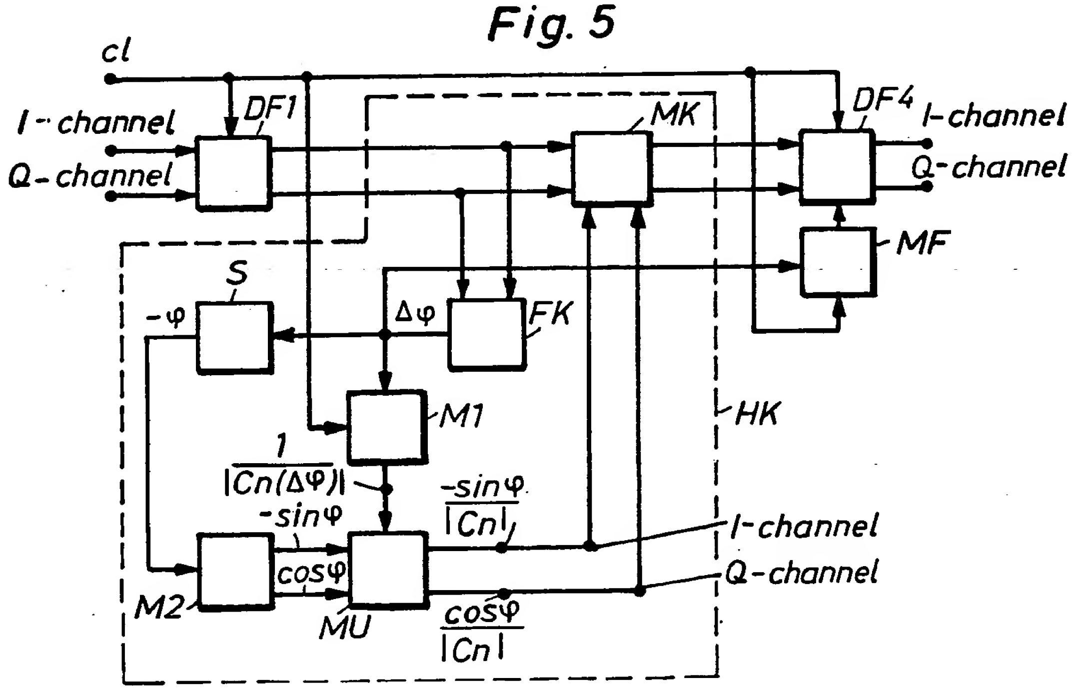


Fig. 5



SPECIFICATION

A radar receiver for pulse Doppler radar equipment

A radar receiver should be able to eliminate such radar echoes, "clutter", caused by reflections from irrelevant targets such as the ground, the sea or precipitation (rain or snow) and only detect the desired moving target, for example an aeroplane. For this purpose, the speed difference of the not desired target relative to the desired target or targets may be used. In a coherent pulse Dopper radar equipment of a known kind, a pulsed high frequency signal at a certain carrier frequency fo is transmitted which, after reflection against a moving target, returns with changed frequency $fo \pm fd$, where the change fd depends on the Doppler shift, i.e. on the radial speed of the moving target relative to the radar equipment. The incoming echo signal is mixed in the radar receiver of the radar equipment with a signal at the carrier frequency fo, a signal at the Doppler frequency fd being obtained. If the transmitted signal (at the carrier frequency fo) were not pulsed, a pure sine wave signal would be obtained whose frequency is the Doppler frequency fd. Since the transmitted signal is pulsed at a pulse frequency $fp = 1/T$, where T is the period time, the receiver will give a pulsed signal which is sine-wave modulated, the modulation having a frequency which is equal to the Doppler frequency fd. Furthermore, the received signal contains frequency components originating from not-desired targets, resulting in that the received signal will not be purely sine-wave modulated. The received signal and the signal from the mixer in the receiver will, consequently, contain a number of desired and not-desired frequency components.

It is already known to provide a filter (a so-called Dopper filter) in a radar receiver for pulse Doppler radar equipment, the task of which is to suppress to as high a degree as possible the frequency components originating from not-desired targets, primarily low frequency components originating from the ground, the sea and precipitation. The Doppler filter could consist of a digital filter which eliminates the components whose frequencies are smaller than a certain value corresponding to a certain target speed. Such a Doppler filter has, within the frequency band determined by the period time T of the radar transmitter, a certain characteristic which is indicated by a broken line in Fig. 1 of the accompanying drawings, it being desirable that the filter have a stop band characteristic for low frequencies, for example smaller than $1/8T$, whereas it has a pass band characteristic for higher frequency values, resulting in that moving targets whose radial speed is greater than the speed of clutter can be detected. The use of a Doppler filter, however, is limited by the magnitude of its pass band. If, for instance, the upper limit frequency $f\ max$ of the filter stop band is approximately $1/8T$ and if the period time T of the radar pulses is limited downwards by the desired range Rmax, it is the case that $T = 2Rmax/c$, where $c =$ the propagation speed, and the fastest clutter speed within the filter stop band will be $vmax = \lambda c/16Rmax$, where $\lambda =$ the radar wavelength. For example, if $\lambda = 1$dm (the S-band) and $Rmax = 10.10^4$m, then $vmax \approx 2$m/s, which means that only the ground clutter can be suppressed by the filter, whereas clutter having higher frequency components will remain uninfluenced.

In the case that the radar equipment operates in the low PRF-mode, that is the period time T is adjusted so that all the radar echoes of interest are reflected and received before the next radar pulse is transmitted, this means that the Doppler frequency fd of the target can be greater than the pulse repetition frequency $1/T$. This in turn means, as appears from Fig. 1, that also the target echo can be suppressed by the Doppler filter for certain so-called "blind speeds", more precisely for the speeds which give Doppler frequencies which are multiples of the frequency $1/T$. It is previously known to avoid suppression of such target echoes by introducing so called "staggering", i.e. the period time T is caused to vary from one pulse to the next of the radar pulses transmitted.

Another known method to eliminate the non-desired clutter spectrum is to carry out a speed compensation before the filtering in the Doppler filter. This means that the clutter speed is estimated, for example by phase measurement during successive sweeps. By, for example, controlling the local oscillator of the receiver, the clutter spectrum can be displaced so that its dominating component assumes the value zero and thus will be situated within the filter stop band. The method, however, assumes that the clutter spectrum has a dominating component which can be easily calculated.

According to the present invention there is provided a receiver for pulse Doppler radar equipment, the receiver including a filter arrangement for reducing not-desired clutter signals within a lower speed range and a higher speed range of a received target echo signal which forms the response of radar pulses transmitted from the equipment with irregular pulse repetition frequency ("staggering"), the arrangement including a first digital filter and a second digital filter of which the frequency limit between a pass band and a stop band of the first filter is such that clutter signals within the lower speed range fall within a stop band of the first filter but the desired target echo signal falls within its pass band, and a circuit connected to an output

of the first filter for calculating a dominant frequency component of clutter signals within the higher speed range and carrying out a transfer in frequency of the clutter signals recovered from the first filter so that the mean frequency of the clutter signals within the higher speed range assumes a value lower than the value before the said transfer, an output of the circuit being connected to an input of the second digital filter, whose stop band substantially coincides with the stop band at low frequencies of the first filter in order to suppress the clutter signals which, before the said transfer, are situated within the higher speed range.

The present invention also comprises Doppler radar equipment including such a receiver.

The invention will now be described by way of example with reference to the accompanying drawings, in which:–

Figure 1 shows a frequency diagram which illustrates on the one hand the frequency spectrum of a radar signal received and on the other hand a certain chosen filter characteristic,

Figure 2 is a block diagram of units included in a radar receiver, which units precede a filter arrangement of the receiver,

Figure 3 is a block diagram for showing the principle of the filter arrangement,

Figure 4 shows details of a digital filter included in the filter arrangement of Fig. 3, and

Figure 5 is a block diagram showing details of the filter arrangement.

In the frequency diagram according to Fig. 1, a clutter frequency spectrum is shown together with the spectrum of an incoming target echo at a certain distance from Doppler radar equipment. The filter characteristic of a digital filter included in a filter arrangement in the receiver of the equipment is indicated by a broken line and has on the one hand a stop band for low frequencies, by way of example for frequencies $<1/8T$, and on the other hand a stop band for frequencies between $7/8T$ and $9/8T$ and therebetween a pass band. The filter characteristic is then periodic with a period equal to $1/T$. The spectrum of the desired target is designated s and the moving clutter has a dominating spectrum sm whose mean frequency is designated *fm*. The Doppler filter arrangement whose design will be closer described in connection with Figs. 4 and 5, has then the task of suppressing on the one hand, the ground clutter spectrum sg and, on the other hand, the spectrum sm of the dominating moving clutter originating primarily from precipitation (rain or snow).

In order to illustrate the signal treatment and the construction of the filter arrangement, the units which precede the filter arrangement will first be described in connection with Fig. 2. At an input A, a signal $A(t) = \cos[2\pi(fo + fd).t + \varphi]$ appears from the duplexer of the radar receiver. The frequency *fd* is the Doppler frequency for the desired target. The signal A(t) is supplied to two channels I and Q which contain mixers B1 and B2 respectively together with analogue-digital converters AD1 and AD2 respectively. To the mixers B1 and B2, reference signals $\cos 2\pi fo.t$ and $\sin 2\pi fo.t$ respectively are supplied from a reference oscillator OSC in the receiver. Output signals are obtained, which are supplied respectively to the analogue-digital converters AD1 and AD2. In these converters, the signals are sampled at sampling instants *tn* by means of clock pulses from a clock circuit CL so that the output signals:

$$XI(tn) = \cos(2\pi fd.tn + \varphi)$$

and

$$XQ(tn) = \sin(2\pi fd.tn + \varphi)$$

are obtained in the channels I and Q respectively. The signals XI(tn) and XQ(tn) can both be represented by the signal:

$$X(tn) = {}_{e}j2\pi fd.tn,\ \varphi = 0$$

The sampling instants can be chosen so that regular sampling is carried out, i.e. $tn = nT$ ($n = 1,2,3,\ldots$) or so that the time between successive sampling pulses varies within a certain time interval NT but the same sampling pattern reappears after the sampling instant $tn = NT$, so called "staggering". In the last mentioned case, it is the case that the sampling is carried out at the instants $\nu Nt + tk$, where $\nu = 0,1 \ldots$ and $k = 0,1,2 \ldots, N-1$.

The Doppler filter arrangement will first be described with reference to Fig. 3. The filter arrangement contains a first digital filter DF1 of a type known per se, suitably a transversal filter which has been designed so that it eliminates ground and sea clutter, i.e. clutter with a low speed (see the filter characteristic according to Fig. 1). Since the filter characteristic of a digital filter is periodic with a period equal to the inverted value of the sampling frequency, the stop band will reappear at frequencies corresponding to certain higher speeds and will reappear periodically if regular sampling is used. In the case that the sampling frequency varies (staggering), the characteristic of the filter DF1 is irregular and a determined position of its stop band cannot be stated except for very low frequencies corresponding to ground and sea clutter.

The filter DF1 can thus, in this latter case, not generally be designed so that clutter with very low speed (from ground or sea) and clutter with higher speed simultaneously can be eliminated. The input signal of the filter is designated X1(tn) and its output signal y1(tn).

To the output of the filter DF1 a circuit HK is connected the construction of which will be described with reference to Fig. 5. The circuit HK carries out a calculation of the clutter which remains after the filtering process in the first filter DF1 and carries out a speed compensation of the mean frequency *fm* of the dominating clutter spectrum. This compensation means that all the frequency components in the incoming signal y1(tn) are moved in frequency so that the remaining moving clutter falls below a certain frequency limit, for example below the value 1/8T in the diagram of Fig. 1. The subsequent digital filter DF2 connected after the circuit HD is designed according to the same principle as the first filter DF1, which is designed so that its stop band coincides with clutter whose frequencies have low values (ground or sea clutter). Thereby, the design problem for the second filter DF2 by using staggering has been transformed to the relatively simple design problem which is the case for the first filter DF1. The filter DF2 thus eliminates the remaining moving clutter (due to precipitation) and the only assumption is that the remaining clutter has a dominating spectrum whose mean frequency *fm* can be calculated in the circuit HK.

Each of filters DF1 and DF2 is a digital filter the design of which is shown in Fig. 4. The filter according to Fig. 4 contains a number of delay circuits, in the example three circuits DL1–DL3 each with a delay T equal to the period time of the radar pulses. The input to delay circuit DL1 and the output of each delay circuit is connected to a respective one of multipliers MU0–MU3 with the coefficients L0(n), L1(n), L2(n) and L3(n) for the filter DF1 and with the coefficients K0(n), K1(n), K2(n) and K3(n) for the filter DF2, where the index (n) indicates that the values of the coefficients can vary for different sampling instants *tn*. All the multiplier outputs are connected to an adder circuit ADD. In the following, only the staggering case is considered, when the time interval between two successive sampling pulses varies in accordance with what has been mentioned above. The case with regular sampling is a special case when $tn = nT$.

The speed compensation of the output signal y1(tn) from the filter DF1 will at first be described as regards signals, and, after that, a suitable embodiment of the circuit HK and the subsequent filter DF2 will be more closely described with reference to Fig. 5.

At staggering, the time between successive sampling pulse varies but the variation is periodic with the period NT, which means that the characteristic of each of the filters DF1 and DF2 is repeated after the instants NT, 2NT, . . . If the input signal to the filter DF1 is $X1(tn) = e^{i2\pi fd.(NT+tn)}$ the output signal from the filter DF1 will be:

$$y1(tn) = \sum_{i=0}^{r1} Li(n).e^{i2\pi fd.(\nu NT + tn - i)}$$

$$= \sum_{i=0}^{r1} Li(n).e^{-i2\pi fd.(nT - tn - i)}.e^{i2\pi fd.(\nu N + n)T}$$

$$= Cn(fd).e^{i2\pi fd.(\nu N + n)T.}$$

where *r*1 is the number of delay circuits in the filter DF1.

In this case, thus, it is the case that the amplitude and the phase of the output signal y1(tn) represented by the factor Cn(fd) is time dependent. The speed compensation means that the signal y1(tn) is divided by the signal $Cn(fm).e^{i2\pi fm.(\nu N + n)T}$ where *fm* is the result of a measurement of the mean frequency of the dominating clutter spectrum after the filter DF1. Thus:

$$X2(tn) = (Cn(fd)/Cn(fm)).e^{i2\pi(fd - fm).(\nu N + n)T}$$

The output signal from the filter DF2 is given from:

$$y2(tn) = \sum_{1=0}^{r2} Kl(n)\frac{Cn-1\ (fd)}{Cn-1\ (fm)}.e^{-j2\pi(fd-fm).1.T}$$
$$.e^{i2\pi(fd-fm).(\nu N + n)T}$$

where *r*2 = the number of delay circuits in the filter DF2. From the expressions for y1(tn), X2(tn) and y2(tn) it appears that (a) if the Doppler frequency $fd \approx 0$, the signal can be eliminated in the first filter DF1, because

$Cn(fd) \approx \sum_{i=0}^{r1} Li(n)$ can be made 0,

(b) if the Doppler frequency $fd \neq 0$ and a correct calculation of this frequency of the dominating clutter spectrum has been carried out in the circuit HK, i.e. $fm \approx fd$, the input signal to the filter DF2 will be

$$X2(tn) = e^{j2\pi(fd - fm).(\nu N + n)T},$$

which represents a low frequency signal and which can be eliminated in the filter DF2 in the same manner as the signal $e^{j2\pi fd.tn}$ was eliminated in the filter DF1 for $fd \approx 0$.

In order to calculate the filter coefficients Li(n) and Kl(n) the following demands are put on the output signals y1(tn) and y2(tn): y1(tn)≈0 when the incoming signal consists of ground clutter, i.e. $fm \approx 0$. Specially, it is demanded that y1(tn) = 0 because $fm = \Delta fK$, where K = 1, . . . r1. ΔfK is chosen within the frequency range of the ground clutter spectrum. This demand can be fulfilled by choosing:

$$Cn(\Delta fK) = 0, \text{ where } K = 1, \ldots r1$$
$$n = 1, \ldots N$$

i.e. the filter DF1 eliminates the ground clutter. The equation Cn(ΔfK) = 0 leads to the following equation system for calculating the filter coefficients Li(n).

$$\sum_{i=0}^{r1} Li(n).e^{-j2\pi\Delta fK.(nT - tn - i)} = 0$$

If ΔfK is chosen symmetrically around the frequency 0, the above indicated relations will lead to real and time dependent coefficients Li(n). The signal y2(tn)≈0 when the input signal consists of moving clutter with the frequency fd. The mean frequency of the clutter has been measured to $fm \approx fd$. Especially, it is demanded that y2(tn) = 0 for $fd - fm = \delta fK$, where K = 1, . . . r2. δfK is then chosen so that $fm + \delta fK$ will be situated within the frequency range for the moving clutter. This condition will give the following equation system:

$$\sum_{1=0}^{r2} K^{1(n)}.\frac{Cn - 1(fm + \delta fK)}{Cn - 1(fm)}.e^{-j2\pi.\delta fK.1.T = 0.}$$

where K = 1, . . . r2.

for determining the filter coefficients K1(n).

Fig. 5 is a block diagram of an embodiment of the circuit HK of Fig. 3 for attaining a speed compensation together with the two filters DF1 and DF2. From the expression for X2(tn) according to the above, it appears that the compensation as regards signals is carried out by dividing the output signal y1(tn) from the filter DF1 by the factor $Cn(fm)_{\cdot e} j2\pi fm\ (\nu N + n)tT$, where fm represents a calculated value of the mean frequency of the dominating spectrum sm of the moving clutter. From the filter DF1, a signal Re{y1(tn)} in the channel I and a signal Im{y1(tn)} in the channel Q is obtained. To each of the channels I and Q, a phase measuring circuit FK is connected for measuring the phase difference between two consecutive and filtered samples. This is carried out by at first measuring the two components of the sample value in the channels I and Q, a value of the phase angle φ1 relative to a certain reference being obtained. After that, and in the same manner, the phase angle φ2 for the next sample value is measured and the difference Δφ = φ2 − φ1 is formed. For each stagger sequence tn, sample values are obtained which give a sequence of phase differences Δφn between two consecutive and filtered samples y1(tn). This sequence Δφn is supplied to an accumulator S for the received phase difference values Δφn during the time period NT corresponding to a complete stagger sequence. The accumulator could be, for example, a feed-back summator.

M1 and M2 designate two memory units, for example PROM's (programmable-read-only-memories). The memory M1 consists of a matrix in which the values of the coefficients Cn(fm) are written for different values of the phase difference Δφ and for the different values of tn in the stagger sequence. For each pair of values tn, Δφ, therefore, a certain value of the coefficient Cn(fm) is obtained, since fm is calculated from the value $\Delta\varphi = 2\pi T.fm$, where T is known.

The memory unit M2 consists of a matrix in the form of a PROM in which the sine and cosine values for different angles φ are listed, which angles are obtained from the accumulator S. The memory unit M1 has only one output, at which the value 1//Cn(fm)/appears, but two inputs, at which the input value Δφ and the clock pulses c1 respectively appear, the latter at the sample instants tn within each interval νNT. The memory unit M2 has a channel I output and a channel Q output at which the values − sin φ and cos φ, respectively, appear, where φ is the

accumulated phase. A multiplier MU is connected to the two outputs of the unit M2 and to the output of the unit M1 for multiplication of the factor 1//Cn(fm)/ by the sine and cosine values respectively of the accumulated phase. Thus, across the channel I and channel Q outputs of the multiplier MU, the two components $-\sin\varphi/|Cn(fm)|$ and $\cos\varphi/|Cn(fm)|$ are obtained which are necessary in order to form the complex value:

$$(1/Cn(fm)).e^{-j2\pi\, fm.(\nu NT+tn)}$$

According to the description above regarding the signal treatment, it is the case that the factor

$$Cn\ (fd)\cdot e^{j2\pi fd.(\nu NT+tn)}$$

should be multiplied by the factor

$$(1/Cn(fm)).e^{-j2\pi fm.(\nu NT+tn)}$$

for the speed compensation. The complex multiplier MK connected to the output of the filter DF1 and to the multiplier MU carries out this complex multiplication, since the channel I and the channel Q components of the complex factors are available as signal values at the respective channels. At the channel I and the channel Q outputs of the multiplier MK, the corresponding signal components of

$$X2(tn) = (Cn(fd)/Cn(fm)).e^{j2\pi(fd-fm).(\nu NT+tn)}$$

will thus be obtained, see the above. The filter DF2 contains a digital transversal filter DF4 of a design according to Fig. 4. In order to attain a good clutter suppression within the entire speed range, it is generally necessary to choose different filter coefficients K1(n) for different measured frequencies fm. The filter coefficients K1(n) are then determined from the above given relation. To the multipliers included in the filter DF4, a memory unit MF is connected, for example in the form of a PROM in which the coefficients K1(n) for each value of $\Delta\varphi$ and for each instant tn are written in matrix form. The memory unit MF is therefore by its two control inputs connected on the one hand to the output of the phase measuring circuit FK at which the value of $\Delta\varphi$ appears and on the other hand to the clock pulse generator, not shown, which generates the clock pulses *c1* in time with the stagger sequence tn (within each interval νNT). The values of the coefficients K1(n) dependent on $\Delta\varphi$ and tn are delivered to the multipliers included in the filter DF4 and at the channels I and Q outputs of the filter the quadrature components of the desired filtered signal y2(tn) are obtained.

The filters DF1 and DF2 are, as mentioned above, designed according to Fig. 4. This figure shows, however, only the design for one channel, for example the channel I, and the multipliers MU0–MU2 multiply the signal components of X1(tn) and X2(tn) in this channel by the corresponding components of the coefficients Li(n) and K1(n). The corresponding filter circuits are present in the other channel Q and at the complex multiplication in the multipliers MU0–MU2, the values in the channels I and Q are mixed together. The filters DF1 and DF2 have for a given order (determined by the number of delay circuits DL1–DL3) a given pass band whose width in known manner can be extended by choosing filters of higher order.

CLAIMS

1. A receiver for pulse Doppler radar equipment, the receiver including a filter arrangement for reducing not-desired clutter signals within a lower speed range and a higher speed range of a received target echo signal which forms the response of radar pulses transmitted from the equipment with irregular pulse repetition frequency ("staggering"), the arrangement including a first digital filter and a second digital filter of which the frequency limit between a pass band and a stop band of the first filter is such that clutter signals within the lower speed range fall within a stop band of the first filter but the desired target echo signal falls within its pass band, and a circuit connected to an output of the first filter for calculating a dominating frequency component of the clutter signals within the higher speed range and carrying out a transfer in frequency of the clutter signals recovered from the first filter so that the mean frequency of the clutter signals within the higher speed range assumes a value lower than the value before the said transfer, an output of the circuit being connected to an input of the second digital filter, whose stop band substantially coincides with the stop band at low frequencies of the first filter in order to suppress the clutter signals which, before the said transfer, are situated within the higher speed range.

2. A receiver according to claim 1, wherein the said circuit includes a phase measuring circuit for determining a sequence of phase differences between two consecutive sampled values filtered in the first filter for each sequence of transmitted radar pulses, a first memory unit for

forming the inverted values of the coefficients which are equal to the sample values of the output signal from the first filter corresponding to a certain phase difference, a multiplier circuit connected between the first and second filter and to the said memory unit for multiplying the output signal with the said inverted value, and a second memory unit connected to the said phase measuring circuit and to the second digital filter in order to store the coefficient values belonging to the second filter for each measured phase difference and for each sample instant within a stagger sequence.

3. A receiver according to claim 2, wherein the said multiplier circuit comprises a complex multiplier having two input pairs of which each pair corresponds to the I and Q channels of the radar receiver, the first input pair being connected to the output of the first filter, an accumulator is connected to the output of the phase measuring circuit to form a mean value of the said sequence of phase differences, a third memory unit being provided to form the sine and the cosine values of the said mean value, and a further multiplier is connected to the said first memory unit and to the said third memory unit to multiply the said sine and cosine values with the said inverted value, the multiplied values being supplied to the second input pair of the complex multiplier across the associated I and Q channels.

4. A radar receiver for a pulse Doppler equipment, substantially as herein described with reference to the accompanying drawings.

5. Pulse Doppler radar equipment including a radar receiver according to any preceding claim.

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